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# (54) ONE-BIT ANALOG-TO-DIGITAL CONVERTERS AND DIGITAL-TO-ANALOG CONVERTERS USING AN ADAPTIVE FILTER HAVING TWO REGIMES OF OPERATION

EIN-BIT ANALOG/DIGITAL-WANDLER UND DIGITAL/ANALOG-WANDLER MIT EINEM ADAPTIVEN FILTER MIT ZWEI BETRIEBSARTEN

CONVERTISSEURS ANALOGIQUES-NUMERIQUES ET CONVERTISSEURS NUMERIQUES-ANALOGIQUES A UN BIT UTILISANT UN FILTRE ADAPTATIF A DEUX REGIMES DE FONCTIONNEMENT

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## Description

### *Background of the Invention*

The invention relates to one-bit analog-to-digital converters (ADCs) and digital-to-analog converters (DACs). More particularly, the invention relates to such converters which employ an adaptive filter having two regimes or modes of operation.

One-bit digital systems have the great virtue that they do not require high precision components. The digital-to-analog converter (DAC) of a one-bit system consists simply of a low-pass filter 2 such as shown in Figure la. The analog-to-digital converter (ADC), such as shown in Figure 1b, generally consists of an identical filter 2, known as the local decoder, a comparator 4 comparing the analog input signal with the output of the local decoder, and a sampling system such as sampler and one-bit quantizer 6, typically implemented as a D flip-flop, clocked at regular intervals to deliver data pulses whose polarity (1 or 0) depends on whether the local decoder's output is positive or negative with respect to the input signal.

Because every ADC contains a DAC (a local decoder - usually just a filter), the following discussion is directed primarily to the filter itself. It will be understood that the filter is employed in an ADC or a DAC.

In the digital-to-analog converters of non-adapting one-bit systems, impulses are fed via a filter to yield the analog output signal. The impulses are of two amplitudes, corresponding to the 1 and 0 of a data bit (e.g., they might be +5 and -5 volts). The nature of the filter determines the type of one-bit system.

In Delta Modulation systems the filter is an integrator or a low-pass filter with a cut-off frequency below or near the bottom of the message band (the range of frequencies occupied by the signal to be conveyed), so that over that band the response of the filter falls progressively with increasing frequency, most commonly at 6 dB/octave.

In Delta-Sigma Modulation systems the filter is a low-pass filter with a cut-off frequency above or near the top of the message band, so that over that band the filter response is substantially flat. For example, in a system designed for high-quality audio where the message band might be 30 Hz to 15 kHz, a delta modulation system might use a single-pole filter with a cut-off at 100 Hz (often called a leaky integrator), while the filter of a delta-sigma modulation system might cut-off at 10 kHz.

One-bit systems may be adaptive. Presently known methods can be broadly divided into two types: amplitude variation and filter adaptation.

Amplitude variation is the method used in conventional adaptive delta modulation. The amplitude of the impulses is varied, either continuously or in discrete steps, before filtering. Control circuitry is designed so that increasing signal amplitude leads to increasing impulse amplitude. In the example of Figure 2a, the size of the input pulses is modulated by a voltage-controlled amplifier (VCA) 8 which precedes a fixed low-pass filter 10, but other methods are possible. An alternative but equivalent configuration, shown in Figure 2b, uses fixed amplitude impulses but follows the filter 10 by the VCA 8. Either approach yields a much improved signal-to-noise ratio for low-level signals, compared with a non-adapting system. The noise has a substantially constant spectrum but varies with signal level, being directly proportional to the pulse amplitude. This variation may be audible as noise modulation. It is particularly likely to be audible in the presence of high-frequency signals, which may mask the rise in noise at high frequencies but not the accompanying rise at low and middle frequencies.

The filter adaptation approach employs fixed amplitude pulses but varies the cut-off frequency of the filter, again either continuously or discretely. Figure 3 shows such an arrangement having a variable low-pass filter 2'. Under no-signal or low-level conditions, the system is conventional delta modulation, generally with a filter frequency substantially below the bottom of the message band. As the signal level increases, the filter moves up but initially it still has a cut-off frequency below the message band. However, once the cut-off frequency is within the message band, the gain below the cut-off frequency becomes constant. Thus the system yields a noise spectrum which is variable, with much less low-frequency noise in the presence of high-frequency signals. At high signal levels the filter slides upwards in frequency to approach the top of the message band; the system changes from delta to delta-sigma modulation.

This technique of adapting the filter gives a better subjective signal-to-noise ratio in the presence of high-level high-frequency signals because under these conditions the low frequency gain of the filter is lower than that of an integrator, and therefore the noise spectrum contains less energy at low frequencies where there is little or no masking.

However, under no-signal conditions the filter is effectively a pure integrator over a wide range of frequencies from below to the top of the message band and beyond. Thus, compared with a system employing amplitude adaptation there is a greater amplification at very low frequencies and therefore a greater proportion of low frequency noise at the bottom of and below the audio spectrum.

This increased low-frequency noise may not of itself be a significant disadvantage since under these signal conditions the absolute filter gain in the audio band is small so the absolute noise level is low. In addition the human ear is very insensitive to low-level low-frequency sounds, so the extra low-frequency noise is unlikely to be audible.

However, the required voltage-controlled filter (VCF) will in practice employ some form of VCA. It is an undesirable property of most designs of VCA that a small proportion of the control signal of added to the controlled signal, so that even when trimmed to minimize

the effect the output of the VCF may contain a voltage or current offset which varies with the position of the filter cut-off frequency. If the variable filter is effectively an integrator under small-signal conditions, this variable offset may be amplified excessively, yielding an audible "thump" on high-frequency transients and a visible displacement of the base-line when the output waveform is observed on an oscilloscope.

As explained above, compared with conventional (amplitude-) adaptive delta modulation, filter adaptation gives reduced audible noise modulation but demands better performance from the VCA used in the adapting circuit. In accordance with the present invention, an ADC or DAC employs an adapting filter which retains the advantage of the sliding low-pass filter under high-level conditions but does not demand less offset from the VCA under low-level conditions.

In addition, in accordance with the present invention, the advantage of the sliding band (variable filter cutoff frequency) is realized under high-level high-frequency conditions but without the excessive very low frequency gain under no-signal conditions, thus reducing very low-frequency noise in the quiescent or low-level state. Thus, ADCs and DACs in accordance with the invention provide the advantages of amplitude- and filter-adapting arrangements while avoiding their weak points.

***Summary of the Invention***

To understand better the advantages of ADCs and DACs according to the present invention, particularly with respect to the adapting filter of such converters, it is useful to explain the behavior of the prior-art filters in more detail.

Consider the bitstream from a one-bit ADC, fully modulated with a sine wave whose frequency is swept across the whole audio spectrum. The modulation consists of a variation in the relative proportions of 1s and 0s, and full modulation means that at the maximum positive excursions of the modulating sine wave almost all bits are of one kind, say 1s, and at the maximum negative excursions almost all bits are of the other kind, 0s. At the zero crossings of the modulating sine wave, 1s and 0s occur with equal frequency. If such a bitstream is fed into a low-pass filter with a cut-off frequency at the top of the audio range, the output will be the swept sine wave with a constant amplitude. Such a filter constitutes a non-adapting delta-sigma DAC.

Consider the required behavior of an adaptive delta demodulator fed with the same bitstream, assuming that the output amplitude is again required to be constant with frequency.

If the adaptation is purely in amplitude (i.e., the size of the bits fed into a leaky integrator is variable, as in Figure 2a), the transfer function from the input data to the analog output is a family of curves all of the same shape, a low-pass filter with a fixed cut-off near the bottom of the message band. To obtain a constant output from the fully modulated data, the gain must rise with increasing frequency (see Figure 4). Curves a1, a2 and a3 show the required function to yield the same output at 100 Hz, 1 kHz and 10 kHz, respectively. Note that the gain, and therefore the noise, at say 100 Hz is much higher when the response has adapted to give the required output at 10 kHz. Curve a0 shows how the response might be when there is no modulation. The gain has fallen dramatically and hence the output noise is very small.

If the adaptation involves a sliding filter as in Figure 3, curves b1, b2 and b3 of Figure 5 show the required response to yield the same output at the same three frequencies, 100 Hz, 1 kHz and 10 kHz, respectively. Even with the high-level high-frequency signal, the gain at low frequencies does not rise more than a few dB (about 10 dB in the example shown) above that needed to reproduce low and middle frequencies, so the low-frequency noise does not rise excessively. Comparing curve b3 with a3, it can be seen that in the presence of 10 kHz the low-frequency gain and hence noise is about 20 dB lower. However, under no-signal conditions, the filter must slide down, for example to curve b0, in order to obtain low noise. Note that although the gain is now low in the middle of the audio band, where noise is most audible, it is much higher at the bottom of and below the message band; comparing curves b0 and a0, the gain at 1 kHz are the same, but the sliding filter has about 12 dB more gain at 20 Hz, and the difference continues to increase at frequencies below the audio band. This excess gain leads to higher noise at very low frequencies, not usually a problem in itself, and to undesired amplification of any variable offset in the variable filter circuit.

Figure 6 shows a different family of adapting curves. The no-signal and "100 Hz" curves, c0 and c1, are identical with those of the amplitude adapter (Figure 4, a0 and a1), but for further adaptation the system "slides" as in the curves of Figure 5 (c2 = b2, c3 = b3)(comparing curves "c" in Figure 6 to curves "b" in Figure 5). It will be apparent from Figure 6 that the advantage of the sliding band (variable filter cutoff frequency) is realized under high-level high-frequency conditions but without the excessive very low frequency gain under no-signal conditions, thus reducing very low-frequency noise in the quiescent or low-level state. The characteristics shown in Figure 6 thus provide the advantages of the amplitude- and filter-adapting arrangements while avoiding their weak points.

The differences in the adaptation approaches can be summarized as follows. In the variable amplitude arrangements, as explained by Figures 2a, 2b and 4, each curve in the family is reached by displacement along the gain axis, i.e., a vertical movement. In the variable filter corner frequency (sliding band) arrangement, as explained by Figures 3 and 5, each curve in the family is reached by displacement along the frequency axis, i.e., a horizontal movement. In an adaptive filter arrange-

ment according to the present arrangement (Figure 6 and the embodiments of Figures 8-11), the curves move initially up the gain axis (vertically) and subsequently along the frequency axis (horizontally)

Figure 7 shows the block diagram of a typical embodiment of a prior art adapting filter. A forward path contains a VCA 12 and an integrator 14 in tandem, with overall negative feedback imposing a maximum forward. gain. The VCA and integrator may occur in either order in Figure 7 and in the embodiments of Figures 8-11, but practical considerations of noise and distortion normally require the order shown. The integrator has a transfer characteristic $\frac{1}{sT}$.

A control signal applied to VCA 12 varies its gain. The control signal is generally a function of the amplitude of the analog input signal to the ADC or the analog output signal of the DAC. As the amplitude of the analog signal increases, the control signal increases to increase the gain of the VCA. Circuits for generating control signals are well-known in the art.

A combiner 16 additively combines the input signal having amplification " ***a*** " with a signal at the output of the negative feedback path having amplification " ***-b*** ". By standard analysis of the feedback system, the overall gain represented as an expression in Laplace transform notation is

$$\frac{a}{b} \cdot \frac{1}{1+\frac{sT}{bc}}$$

This expression represents a low-pass filter with a constant gain in its passband of $\frac{a}{b}$ and a cut-off frequency $\frac{bc}{2\pi T}$, which varies with the VCA gain ***c***.

While the prior art has been explained above on the basis of an integrator, this explanation also applies to the case where a low-pass filter is substituted for the integrator provided the cut-off frequency of the low-pass filter is below the audible range. In fact, a "true" integrator has infinite gain at zero frequency and is therefore unrealizable. The document IEEE Transactions on Consumer Electronics, Nov.1989, vol.35, No.4, New York, US, pages 767-773; T.ISHIKAWA et al.: "One-bit A/D D/A Converters IC for Audio Delay" discloses a variable filter similar to that shown in Fig. 8 wherein the cut-off frequency, however, is at about 1 Hz, i.e. far below the audible range. Thus, for all frequencies in the audible range the low-pass filter of this prior art behaves as an integrator.

## Brief Description of the Drawings

Figure 1a is a block diagram of a one-bit digital-to-analog converter (ADC).

Figure 1b is a block diagram of a one-bit analog-to-digital converter (DAC).

Figure 2a is a block diagram of one type of amplitude-adaptive delta demodulator.

Figure 2b is a block diagram of another type of amplitude-adaptive delta demodulator, equivalent to the arrangement of Figure 2a.

Figure 3 is a block diagram of a frequency-adaptive delta demodulator.

Figure 4 is a series of waveforms showing theoretical transfer functions of a one-bit DAC employing a variable amplitude filter.

Figure 5 is a series of waveforms showing theoretical transfer functions of a one-bit DAC employing a variable frequency (sliding band) filter.

Figure 6 is a series of waveforms showing theoretical transfer functions of a one-bit DAC employing an adaptive filter in accordance with the present invention.

Figure 7 is a block diagram of a prior art DAC using a variable low-pass filter.

Figure 8 is a block diagram of an adaptive filter for use in an ADC or a DAC in accordance with the present invention.

Figure 9 is a more detailed block diagram of an adaptive filter for use in an ADC or a DAC in accordance with the present invention.

Figure 10 is a block diagram of an alternative embodiment in which an arrangement substantially equivalent to an adaptive filter for use in an ADC or a DAC in accordance with the present invention is shown.

Figure 11 is a more detailed block diagram of an alternative embodiment in which an arrangement equivalent to an adaptive filter for use in an ADC or a DAC in accordance with the present invention is shown.

## Detailed Description of the Preferred Embodiments

Referring now to Figure 8, a variable filter for use in an ADC or a DAC in accordance with the present invention is shown. As mentioned above, such a filter is used in an ADC and a complementary DAC in the manner of Figures 1a and 1b. In the embodiment of Figure 8, the integrator of Figure 7 is replaced by a fixed low-pass filter 18 having a transfer characteristic $\frac{1}{1+sT}$ with the cut-off frequency within the audible range. As in the Figure 7 arrangement a control signal applied to VCA 12 varies its gain. The comments regarding the control signal in Figure 7 also apply here. The overall gain of this circuit is

$$\frac{ac}{1+bc} \cdot \frac{1}{1+\frac{sT}{1+bc}}.$$

When the VCA gain is low so that $bc << 1$, this simplifies to

$$ac \cdot \frac{1}{1+sT}.$$

This expression is that of a low-pass filter at a fixed frequency $\frac{1}{2\pi T}$ having a gain in its passband of ***ac***, i.e.,

variable and proportional to the VCA gain.

When the VCA gain is high so that $bc >> 1$, the full expression reduces to

$$\frac{a}{b} \cdot \frac{1}{1+\frac{sT}{bc}}.$$

This is the same as the case for the integrator arrangement of Figure 7; a sliding filter with a fixed passband gain of $\frac{a}{b}$ and a variable cut-off frequency $\frac{bc}{2\pi T}$.

Hence, the configuration of Figure 8 achieves the desired variable response, exhibiting as an ideal circuit, pure gain change for low values of $c$ and pure sliding band for high values of $c$. The transition from one regime or mode of operation to the other occurs at a threshold where $bc = 1$.

A preferred embodiment of the adaptive filter for use in ADCs and DACs of the present invention is shown in more detail in Figure 9. As is the embodiment of Figure 8, such a filter is used in an ADC and a complementary DAC in the manner of Figures 1a and 1b. The Figure 9 embodiment uses a variable transconductance amplifier 20 (e.g., a National Semiconductor IC type LM13700) in series with an operational amplifier ("op. amp") 22. In the manner of Figure 7 a control signal is applied to amplifier 20 to vary its gain. The comments regarding the control signal in Figure 7 also apply here. The op. amp has a local feedback path with a capacitor C1 and a resistor R3 in parallel. The variable transconductance amplifier 20 functions in the manner of VCA 12 in Figure 8 and its amplification is also varied by an control signal. The op. amp 22 and its local feedback path function in the manner of low-pass filter 18 in Figure 8. Resistor R2 provides the negative feedback path to the node 24 at the positive input of amplifier 20. The input, fed through resistor R1, is summed with the output of the negative feedback path at the node 24.

The arrangement of Figure 9 differs from the prior art in the addition of resistor R3, which converts the operational amplifier ("op. amp") from an integrator (in the absence of R3) to a fixed low-pass filter. This resistor lowers the impedance of the local feedback around the op. amp. at low frequencies (where C1 presents a high reactance), and hence reduces the amplification of any offset current emerging from the variable transconductance. The circuit of Figure 9 has an overall transfer characteristic

$$-g_m \cdot \frac{R_2R_3}{R_1+R_2} \cdot \frac{1}{1+g_m \cdot \frac{R_1R_3}{R_1+R_2} + j\omega R_3C_1}.$$

where $g_m$ is the variable transconductance, proportional to the control signal. When $g_m$ is low, this simplifies to

$$-g_m \cdot \frac{R_2R_3}{R_1+R_2} \cdot \frac{1}{1+j\omega R_3C_1},$$

which represents a fixed low-pass filter with a variable passband gain. When $g_m$ is high, the expression simplifies to

$$-\frac{R_2}{R_1} \cdot \frac{1}{1+j\omega\frac{(R_1+R_2)C_1}{g_m R_1}},$$

which represents a low-pass filter of fixed gain but variable frequency. The transition or threshold between the two regimes or modes of operation occurs where

$$g_m = \frac{R_1+R_2}{R_1R_3}.$$

An alternative preferred embodiment of a variable filter for use in an ADC or a DAC in accordance with the present invention is shown in Figure 10. This alternative embodiment also provides two regimes of operation and also reduces the effect of variable offset resulting from shortcomings in VCAs. In this arrangement the variable filter characteristic is achieved by placing a "pre-emphasis" type network 26 in series with a VCA 12' in the feedback path instead of providing a low-pass filter in series with a VCA in the forward path as in the embodiment of Figure 8. The embodiment of Figure 8 is preferred inasmuch as the Figure 10 embodiment is less convenient to implement, requiring additional amplifiers in order to provide a voltage output from VCA 12' rather than, as in Figure 8, a current output which is normally provided by a VCA.

As in the previous embodiments, a control signal applied to the VCA varies its gain. However, the control signal acts in the opposite sense to that of the Figure 8 embodiment. The comments regarding the control signal in Figure 7 also apply here. Network 26 has a transfer characteristic $1 + sT$, which is the reciprocal of the Figure 8 network 18 characteristic. The network 26 transfer characteristic is the sum of a fixed unity gain and a differentiator. Such a network is widely used to provide pre-emphasis in FM radio (where $T$ is 75 µs in the United States) and is easily realizable.

A combiner 16 additively combines the input signal having amplification "$+\frac{a}{b}$" with a signal at the output of the negative feedback path having amplification "$-\frac{1}{b}$". VCA 12' has a gain of $\frac{1}{c}$. The overall gain of the circuit is the same as that of the Figure 8 embodiment, namely

$$\frac{ac}{1+bc} \cdot \frac{1}{1+\frac{sT}{1+bc}}.$$

As in the Figure 8 embodiment, when $bc<<1$, this simplifies to

$$ac \cdot \frac{1}{1+sT},$$

however, this occurs when the VCA gain is high rather than low.

When the VCA gain is low so that $bc>>1$, the full expression reduces to

$$\frac{a}{b} \cdot \frac{1}{1+\frac{sT}{bc}}.$$

This is the same as the case for the integrator arrangement of Figure 7; a sliding filter with a fixed passband gain of $\frac{a}{b}$ and a variable cut-off frequency $\frac{bc}{2\pi T}$.

Hence, the configuration of Figure 9 also achieves the desired variable response, exhibiting as an ideal circuit, pure gain change for low values of $\frac{1}{c}$ (rather than $c$) and pure sliding band for high values of $\frac{1}{c}$ (rather than $c$). The transition from one regime or mode of operation to the other occurs at a threshold where $bc = 1$. Because the control circuit acts in the opposite sense from that of the Figure 8 embodiment, when the control signal is above a threshold the characteristic is that of a low-pass filter with a fixed cut-off frequency but with variable passband gain and when the control signal has a value below the threshold the characteristic is that of a low-pass filter with a variable cut-off frequency but a fixed passband gain.

The Figure 10 alternative preferred embodiment of a variable filter for use in an ADC or a DAC in accordance with the present invention is shown in more detail Figure 11. A variable transconductance amplifier 28 (e. g., a National Semiconductor IC type LM13700) is located in series with an op. amp 30 in the negative feedback loop of op. amp 32. In the manner of the previously described embodiments, a control signal is applied to amplifier 28 to vary its gain. The comments regarding the control signal in Figure 7 also apply here. The op. amp 30 has a local input and feedback path with a capacitor C2 and a resistor R6 in parallel and a resistor R7 in series to provide the pre-emphasis function. Op. amp 32 has a local feedback path with resistor R5. The input to the overall arrangement is fed through resistor R4 and additively combined at a summing node 34 with the output of the negative feedback path from the amplifier 28.

The circuit has an overall transfer characteristic

$$-\frac{R_5}{R_4} \cdot \frac{1}{1 + g_m \cdot \frac{R_5 R_7}{R_6}(1+j\omega R_6 C_2)},$$

where $g_m$ is the variable transconductance, proportional to the control signal. When $g_m \cdot \frac{R_5 R_7}{R_6} >>1$, this simplifies to

$$-\frac{1}{g_m} \cdot \frac{R_6}{R_4 R_7} \cdot \frac{1}{1+j\omega R_6 C_2},$$

which represents a fixed low-pass filter with a passband gain which varies inversely with $g_m$. When $g_m \cdot \frac{R_5 R_7}{R_6} <<1$, the expression simplifies to

$$-\frac{R_5}{R_4} \cdot \frac{1}{1+j\omega g_m R_5 R_7 C_2},$$

which represents a low-pass filter of fixed gain but variable frequency. The transition or threshold between the two regimes or modes of operation occurs where

$$g_m \cdot \frac{R_5 R_7}{R_6} = 1.$$

The invention may be implemented using circuit arrangements and topologies other than those specifically disclosed. In addition, although analog embodiments are disclosed, the invention may be implemented either wholly or partially in the digital domain. Although a purely digital implementation would not be subject to the VCA offset current problem which is solved by analog embodiments of the invention, digital implementations share with analog implementations the advantages of the two regimes of operation and the reduction in very low-frequency noise in the quiescent or low-level state.

## Claims

1. An adaptive one-bit audio digital-to-analog converter comprising

   an input for receiving data pulses,
   adaptive low-pass filter means (12, 16, 18; 12', 16, 26), coupled to said input, having under low audio level conditions one regime of operation in which the characteristic is that of a low-pass filter with a fixed cut-off frequency but with variable passband gain, and under high audio level conditions another regime of operation in which the characteristic is that of a low-pass filter with a variable cut-off frequency but a fixed passband gain, and
   an output coupled to receive the output of said adaptive low-pass filter means.

2. An adaptive one-bit audio analog-to-digital converter comprising

   local decoder means, said local decoder

means including adaptive low-pass filter means (12, 16, 18; 12', 16, 26) having under low audio level conditions one regime of operation in which the characteristic is that of a low-pass filter with a fixed cut-off frequency but with variable passband gain, and under high audio level conditions another regime of operation in which the characteristic is that of a low-pass filter with a variable cut-off frequency but a fixed passband gain, and

means (4, 6) for comparing and sampling input analog signals with respect to signals from said local decoder means to provide output data pulses having a polarity, 1 or 0, depending on whether the local decoder signal is positive or negative with respect to the input signal,

said output data pulses being applied to said local decoder means.

3. Apparatus according to claim 1 or claim 2 wherein the filter characteristic of said adaptive low-pass filter means (12, 16, 18; 12', 16, 26) varies in response to a control signal such that when the control signal has values on one side of a threshold the characteristic is that of a low-pass filter with a fixed cut-off frequency but with variable passband gain and when the control signal has values on the other side of the threshold the characteristic is that of a low-pass filter with a variable cut-off frequency but a fixed passband gain.

4. Apparatus according to claim 3 wherein said adaptive low-pass filter means includes the

combining means (16; R1, 24; R4, 34) receiving the input data pulses applied to the converter and signals from a negative feedback path for additively combining the input signals and the negative feedback path signals, and

signal path means including

a forward path receiving the output of said combining means and providing at its output the output of the converter and the input to the negative feedback path, and

the negative feedback path (R2; 30, R6, C2, R7, 28) coupling the output of said forward path to said combining means,

said signal path means further including a variable transconductance amplifier (12; 20; 28) in which the unwanted offset of the variable transconductance amplifier is reduced as the variable gain of said adaptive low-pass filter means decreases.

5. Apparatus according to claim 4 in which said signal path means further includes an operational amplifier (22) in series with said variable transconductance amplifier (12; 20), said operational amplifier and variable transconductance amplifier located in said forward path.

6. Apparatus according to claim 4 in which said signal path means further includes an operational amplifier (30) in series with said variable transconductance amplifier (28), said operational amplifier and variable transconductance amplifier located in said negative feedback path.

7. Apparatus according to claim 5 wherein said operational amplifier (22) has resistance (R3) and capacitance (C1) in parallel in its feedback path, such that the resistance lowers the impedance of the feedback path of the operational amplifier at low frequencies and, hence, reduces the amplification of any offset current produced by the variable transconductance amplifier (20).

8. Apparatus according to claim 7 wherein said adaptive low-pass filter (12, 16, 18; 12', 16, 26) has an overall transfer characteristic represented by the expresssion

$$-g_m \cdot \frac{R_2R_3}{R_1+R_2} \cdot \frac{1}{1+g_m \cdot \dfrac{R_1R_3}{R_1+R_2}+j\omega R_3C_1},$$

where $g_m$ is the variable transconductance proportional to the control signal, whereby when $g_m$ is low the expression representing the overal transfer characteristic simplifies to the expression

$$-g_m \cdot \frac{R_2R_3}{R_1+R_2} \cdot \frac{1}{1+j\omega R_3C_1},$$

which represents a fixed low-pass filter with a variable gain, and when $g_m$ is high, the expression representing the overall transfer characteristic simplifies to the expression

$$-\frac{R_2}{R_1} \cdot \frac{1}{1+j\omega \dfrac{(R_1+R_2)C_1}{g_mR_1}},$$

which represents a low-pass filter of fixed gain but variable frequency, and the transition between the two regimes of operation occurs where

$$g_m = \frac{R_1+R_2}{R_1R_3}$$

## Patentansprüche

1. Adaptiver Ein-Bit-Audio-Digital/Analog-Umsetzer, umfassend

   einen Eingang zum Empfang von Datenimpulsen,
   eine adaptive Tiefpaßfilter-Anordnung (12, 16, 18; 12', 16; 26), die mit dem Eingang gekoppelt ist und unter Bedingungen eines niedrigen Audiopegels einen Betriebsbereich aufweist, bei dem die Kennlinie diejenige eines Tiefpaßfilters mit einer festen Grenzfrequenz, aber mit variabler Durchlaßbandverstärkung ist, so wie unter Bedingungen hohen Audiopegels einen anderen Betriebsbereich aufweist, bei dem die Kennlinie diejenige eines Tiefpaßfilters mit variabler Grenzfrequenz, aber fester Durchlaßbandverstärkung ist, und
   einen Ausgang, der so angeschlossen ist, daß er das Ausgangssignal der adaptiven Tiefpaßfilter-Anordnung empfängt.

2. Adaptiver Ein-Bit-Audio-Analog/Digital-Umsetzer, umfassend

   eine lokale Dekoderanordnung, die eine adaptive Tiefpaßfilter-Anordnung (12, 16, 18; 12', 16, 26) enthält, welche unter Bedingungen eines niedrigen Audiopegels einen Betriebsbereich aufweist, bei dem die Kennlinie diejenige eines Tiefpaßfilters mit einer festen Grenzfrequenz, aber variabler Durchlaßbandverstärkung ist, sowie unter Bedingungen hohen Audiopegels einen anderen Betriebsbereich aufweist, bei dem die Kennlinie diejenige eines Tiefpaßfilters mit einer variablen Grenzfrequenz, aber einer festen Durchlaßbandverstärkung ist,
   eine Anordnung (4, 6) zum Vergleichen und Abtasten analoger Eingangssignale im Hinblick auf Signale von der lokalen Dekoderanordnung zur Lieferung von Ausgangsdatenimpulsen mit einer Polarität, 1 oder 0, abhängig davon, ob das lokale Dekodersignal positiv oder negativ bezüglich des Eingangssignals ist,
   wobei die Ausgangsdatenimpulse an die lokale Dekoderanordnung angelegt werden.

3. Vorrichtung nach Anspruch 1 oder 2, bei der die Filterkennlinie der adaptiven Tiefpaßfilter-Anordnung (12, 16, 18; 12', 16, 26) als Antwort auf ein Steuersignal variiert derart, daß, wenn das Steuersignal Werte auf einer Seite eines Schwellenwerts aufweist, die Kennlinie diejenige eines Tiefpaßfilters mit einer festen Grenzfrequenz, aber mit variabler Durchlaßbandverstärkung ist, und, wenn das Steuersignal Werte auf der anderen Seite des Schwellenwerts aufweist, die Kennlinie diejenige eines Tiefpaßfilters mit einer variablen Grenzfrequenz, aber einer festen Durchlaßbandverstärkung ist.

4. Vorrichtung nach Anspruch 3, bei der die adaptive Tiefpaßfilter-Anordnung enthält

   eine Kombiniereinrichtung (16; R1, 24; R4, 34), welche die an den Umsetzer angelegten Eingangsdatenimpulse sowie Signale von einem Gegenkopplungsweg empfängt, um die Eingangssignale und die Gegenkopplungswegsignale additiv zu kombinieren, und
   eine Signalweganordnung, enthaltend
   einen Vorwärtsweg, der das Ausgangssignal der Kombiniereinrichtung empfängt und an seinem Ausgang das Ausgangssignal des Umsetzers und das Eingangssignal für den Gegenkopplungsweg liefert, und
   den Gegenkopplungsweg (R2; 30, R6, C2, R7, 28), der den Ausgang des Vorwärtswegs mit der Kombiniereinrichtung koppelt,
   wobei die Signalweganordnung femer einen variablen Transkonduktanzverstärker (12; 20) enthält, worin der unerwünschte Offset des variablen Transkonduktanzverstärkers verringert wird, wenn die variable Verstärkung der adaptiven Tiefpaßfilteranordnung abnimmt.

5. Vorrichtung nach Anspruch 4, bei der die Signalweganordnung ferner einen Operationsverstärker (22) in Reihe mit dem variablen Transkonduktanzverstärker (12; 20) aufweist, wobei der Operationsverstärker und der variable Transkonduktanzverstärker in dem Vorwärtsweg angeordnet sind.

6. Vorrichtung nach Anspruch 4, bei der die Signalweganordnung ferner einen Operationsverstärker (30) in Reihe mit dem variablen Transkonduktanzverstärker (28) aufweist, wobei der Operationsverstärker und der variable Transkonduktanzverstärker in dem Gegenkopplungsweg angeordnet sind.

7. Vorrichtung nach Anspruch 5, bei der der Operationsverstärker (22) eine Parallelschaltung aus einem Widerstand (R3) und einer Kapazität (C1) in seinem Rückkopplungsweg aufweist derart, daß der Widerstand die Impedanz des Rückkopplungswegs des Operationsverstärkers bei niedrigen Frequenzen verringert und damit die Verstärkung irgendeines von dem variablen Transkonduktanzverstärker (20) erzeugten Offsetstroms verringert.

8. Vorrichtung nach Anspruch 7, bei dem das adaptive Tiefpaßfilter (12, 16, 18; 12', 16, 26) eine Gesamtübertragungskennlinie aufweist, dargestellt durch den Ausdruck

$$-g_m \cdot \frac{R_2R_3}{R_1+R_2} \cdot \frac{1}{1+g_m \cdot \frac{R_1R_3}{R_1+R_2}+j\omega R_3C_1},$$

wobei $g_m$ die variable Transkonduktanz ist, die proportional dem Steuersignal ist, wodurch, wenn $g_m$ niedrig ist, der die Gesamtübertragungskennlinie darstellende Ausdruck sich vereinfacht zu dem Ausdruck

$$-g_m \cdot \frac{R_2R_3}{R_1+R_2} \cdot \frac{1}{1+j\omega R_3C_1},$$

welcher ein festes Tiefpaßfilter mit einer variablen Verstärkung darstellt, während, wenn $g_m$ hoch ist, sich der die Gesamtübertragungskennlinie darstellende Ausdruck vereinfacht zu dem Ausdruck

$$-\frac{R_2}{R_1} \cdot \frac{1}{1+j\omega \frac{(R_1+R_2)C_1}{g_mR_1}},$$

welcher ein Tiefpaßfilter fester Verstärkung, aber variabler Frequenz darstellt, und der Übergang zwischen den beiden Betriebsbereichen tritt auf, wo

$$g_m = \frac{R_1+R_2}{R_1R_3}$$

**Revendications**

1. Convertisseur audio numérique - analogique adaptatif à un bit comprenant

   une entrée pour la réception d'impulsions de données,
   des moyens de filtre passe-bas adaptatifs (12, 16, 18; 12', 16, 26), couplés à ladite entrée, ayant dans des conditions de bas niveau audio un régime de fonctionnement dans lequel la caractéristique est celle d'un filtre passe-bas avec une fréquence de coupure fixe mais avec un gain variable dans la bande passante, et dans des conditions de niveau audio élevé un autre régime de fonctionnement dans lequel la caractéristique est celle d'un filtre passe-bas avec une fréquence de coupure variable mais un gain fixe dans la bande passante, et
   une sortie couplée pour recevoir la sortie desdits moyens de filtre passe-bas adaptatifs.

2. Convertisseur audio analogique - numérique adaptatif à un bit comprenant

   des moyens de décodeur local, lesdits moyens de décodeur local comprenant des moyens de filtre passe-bas adaptatifs (12, 16, 18; 12', 16, 26) ayant dans des conditions de bas niveau audio un régime de fonctionnement dans lequel la caractéristique est celle d'un filtre passe-bas avec une fréquence de coupure fixe mais avec un gain variable dans la bande passante, et dans des conditions de niveau audio élevé un autre régime de fonctionnement dans lequel la caractéristique est celle d'un filtre passe-bas avec une fréquence de coupure variable mais un gain fixe dans la bande passante, et
   des moyens (4, 6) pour comparer et échantillonner des signaux analogiques d'entrée par rapport à des signaux provenant desdits moyens de décodeur local pour fournir des impulsions de données de sortie ayant une polarité, 1 ou 0 suivant que le signal du décodeur local est positif ou négatif par rapport au signal d'entrée,
   lesdites impulsions de données de sortie étant appliquées auxdits moyens de décodeur local.

3. Dispositif selon la revendication 1 ou 2 dans lequel la caractéristique de filtre desdits moyens de filtre passe-bas adaptatifs (12, 16, 18; 12', 16, 26) varie en réponse à un signal de commande de façon que lorsque le signal de commande a des valeurs situées d'un côté d'un seuil la caractéristique soit celle d'un filtre passe-bas avec une fréquence de coupure fixe mais avec un gain variable dans la bande passante et lorsque le signal de commande a des valeurs situées de l'autre côté du seuil la caractéristique soit celle d'un filtre passe-bas avec une fréquence de coupure variable mais un gain fixe dans la bande passante.

4. Dispositif selon la revendication 3 dans lequel lesdits moyens de filtre passe-bas adaptatifs comprennent

   des moyens de combinaison (16; R1, 24; R4, 34) recevant les impulsions de données d'entrée appliquées au convertisseur et des signaux d'un trajet de contre-réaction pour combiner de façon additive les signaux d'entrée et les signaux du trajet de contre-réaction, et
   des moyens de trajet de signal comprenant
   un trajet direct recevant la sortie desdits moyens de combinaison et fournissant à sa sortie la sortie du convertisseur et l'entrée au trajet de contre-réaction, et
   le trajet de contre-réaction (R2; 30, R6, C2, R7, 28) couplant la sortie dudit trajet direct auxdits

moyens de combinaison,
lesdits moyens de trajet de signal comprenant de plus un amplificateur à transconductance variable (12; 20; 28) dans lequel le décalage indésirable de l'amplificateur à transconductance variable est réduit lorsque le gain variable desdits moyens de filtre passe-bas adaptatifs diminue.

5. Dispositif selon la revendication 4 dans lequel lesdits moyens de trajet de signal comprennent de plus un amplificateur opérationnel (22) en série avec ledit amplificateur à transconductance variable (12; 20), lesdits amplificateur opérationnel et amplificateur à transconductance variable étant placés dans ledit trajet direct.

6. Dispositif selon la revendication 4 dans lequel lesdits moyens de trajet de signal comprennent de plus un amplificateur opérationnel (30) en série avec ledit amplificateur à transconductance variable (28), lesdits amplificateur opérationnel et amplificateur à transconductance variable étant placés dans ledit trajet de contre-réaction .

7. Dispositif selon la revendication 5 dans lequel ledit amplificateur opérationnel (22) possède une résistance (R3) et un condensateur (C1) en parallèle dans son trajet de contre-réaction, de façon que la résistance abaisse l'impédance du trajet de contre-réaction de l'amplificateur opérationnel aux fréquences basses et réduise ainsi l'amplification de tout courant de décalage généré par l'amplificateur à transconductance variable (20).

8. Dispositif selon la revendication 7 dans lequel ledit filtre passe-bas adaptatif (12, 16, 18; 12', 16, 26) a une caractéristique de transfert résultante représentée par l'expression

$$-g_m \cdot \frac{R_2R_3}{R_1+R_2} \cdot \frac{1}{1+g_m \cdot \frac{R_1R_3}{R_1+R_2} + j\omega R_3C_1},$$

dans laquelle gm est la transconductance variable proportionnelle au signal de commande, et dans laquelle lorsque gm est faible l'expression représentant la caractéristique de transfert résultante se simplifie en l'expression

$$-g_m \cdot \frac{R_2R_3}{R_1+R_2} \cdot \frac{1}{1+j\omega R_3C_1},$$

qui représente un filtre passe-bas fixe avec un gain variable, et lorsque gm est élevée l'expression représentant la caractéristique de transfert résultante se simplifie en l'expression

$$-\frac{R_2}{R_1} \cdot \frac{1}{1+j\omega \frac{(R_1+R_2)C_1}{g_mR_1}},$$

qui représente un filtre passe-bas de gain fixe mais de fréquence variable, et la transition entre les deux régimes de fonctionnement a lieu lorsque

$$g_m = \frac{R_1+R_2}{R_1R_3}.$$

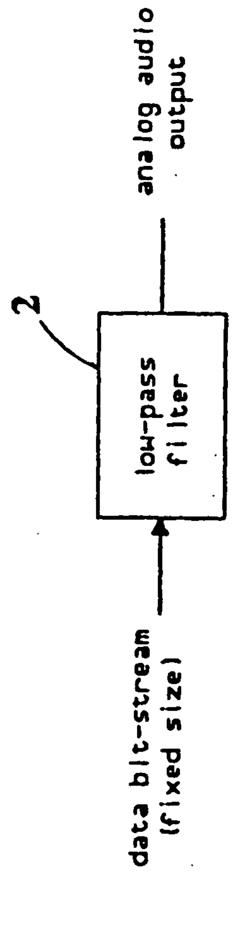


Figure 1a
(Prior Art)

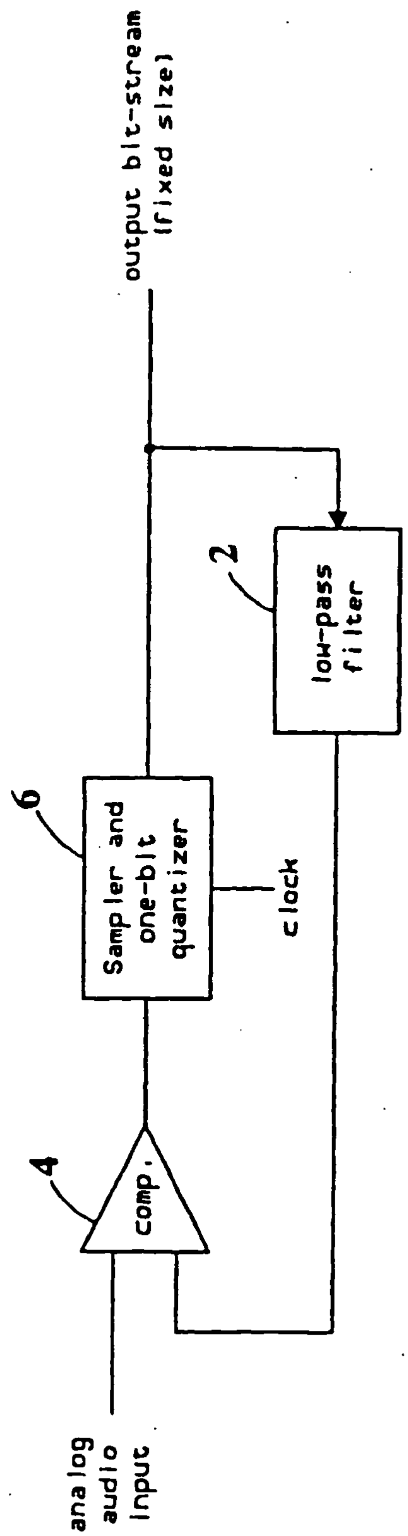


Figure 1b
(Prior Art)

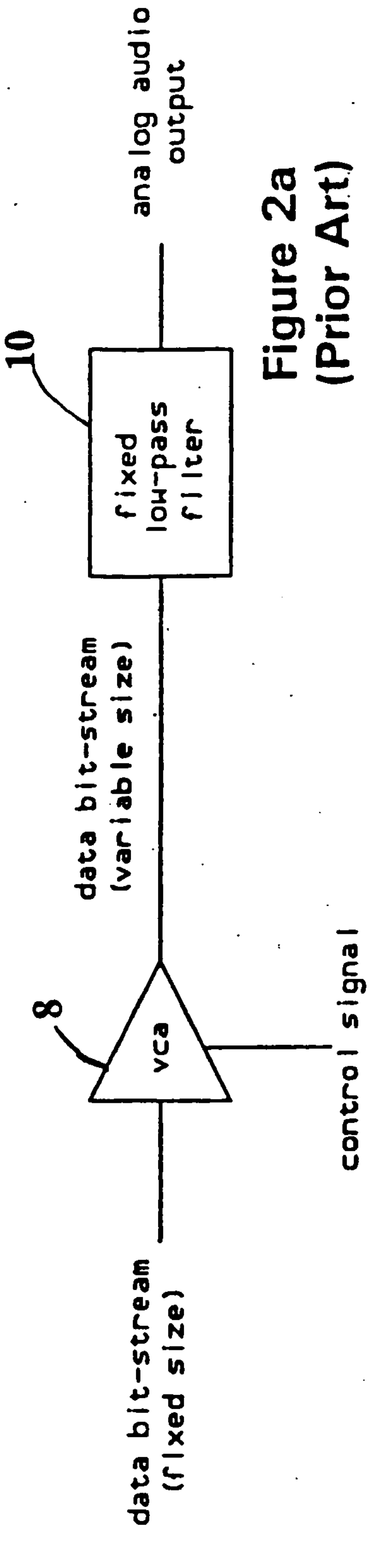


Figure 2a
(Prior Art)

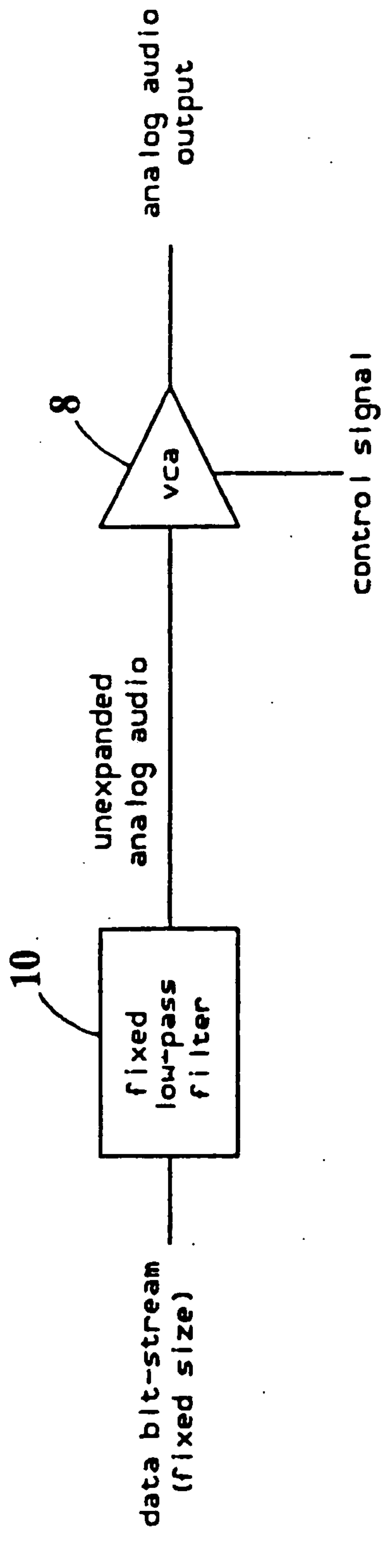


Figure 2b
Prior Art

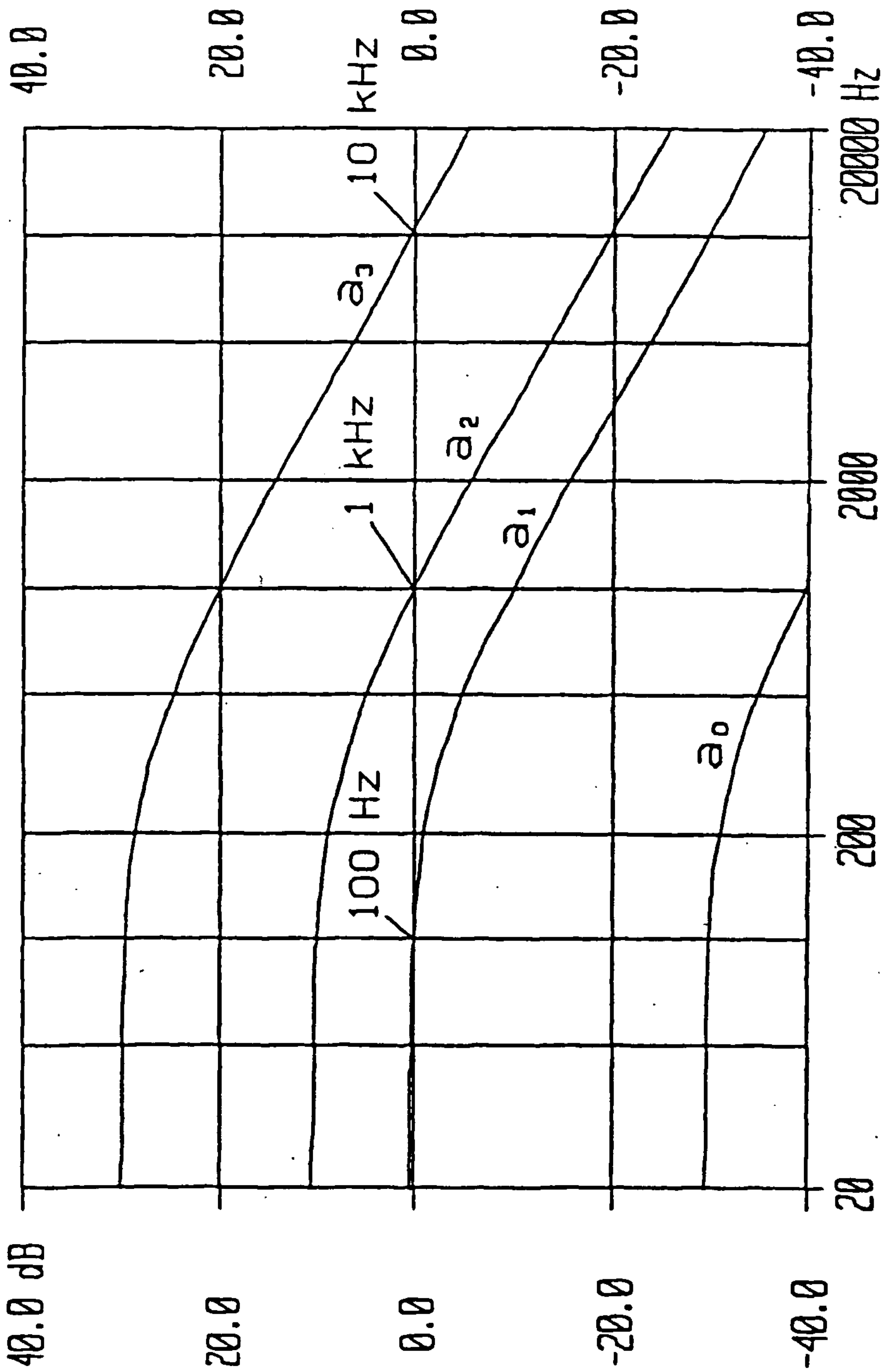


**Figure 4**
**(Prior Art)**

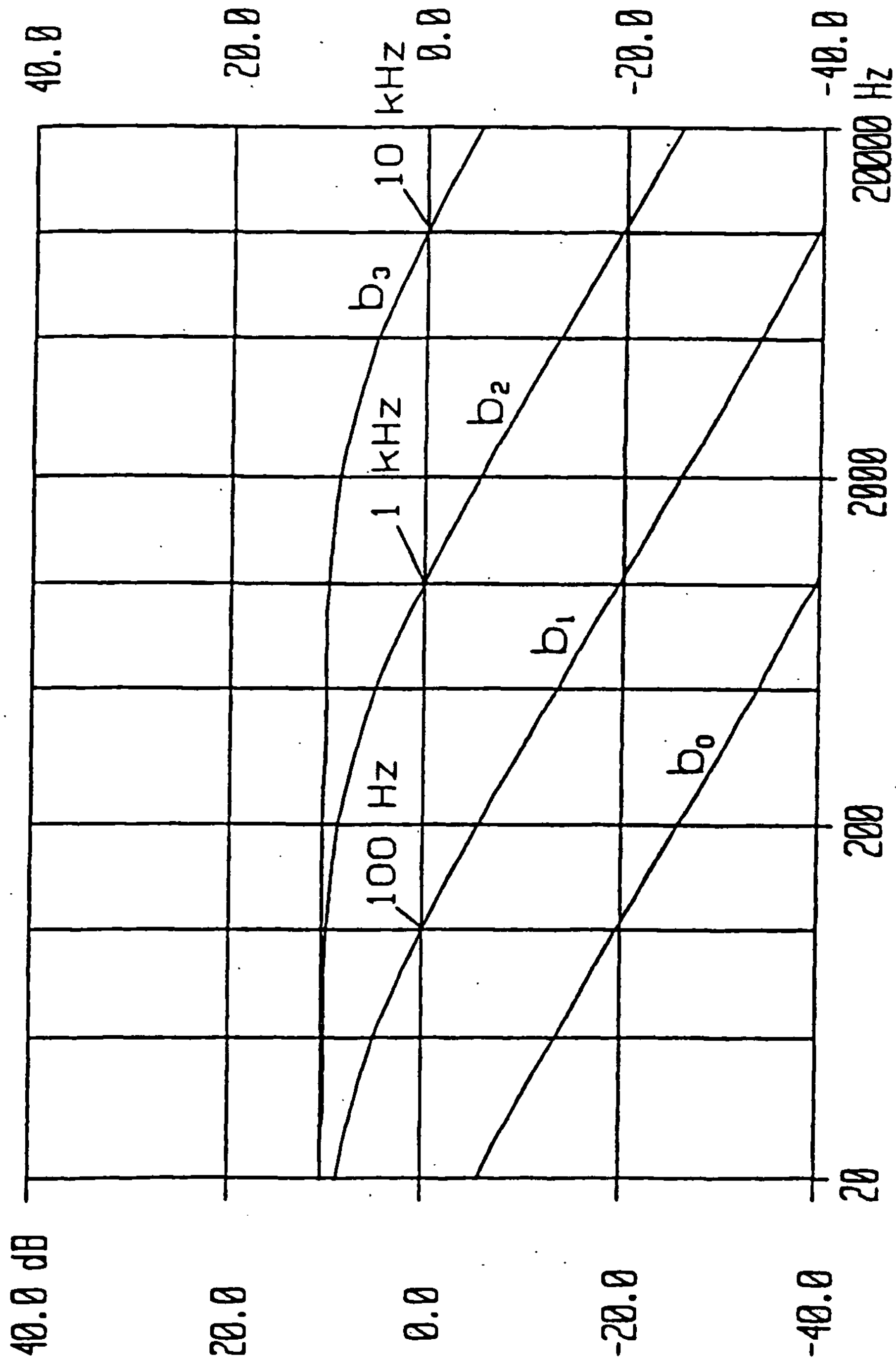


**Figure 5**
**(Prior Art)**

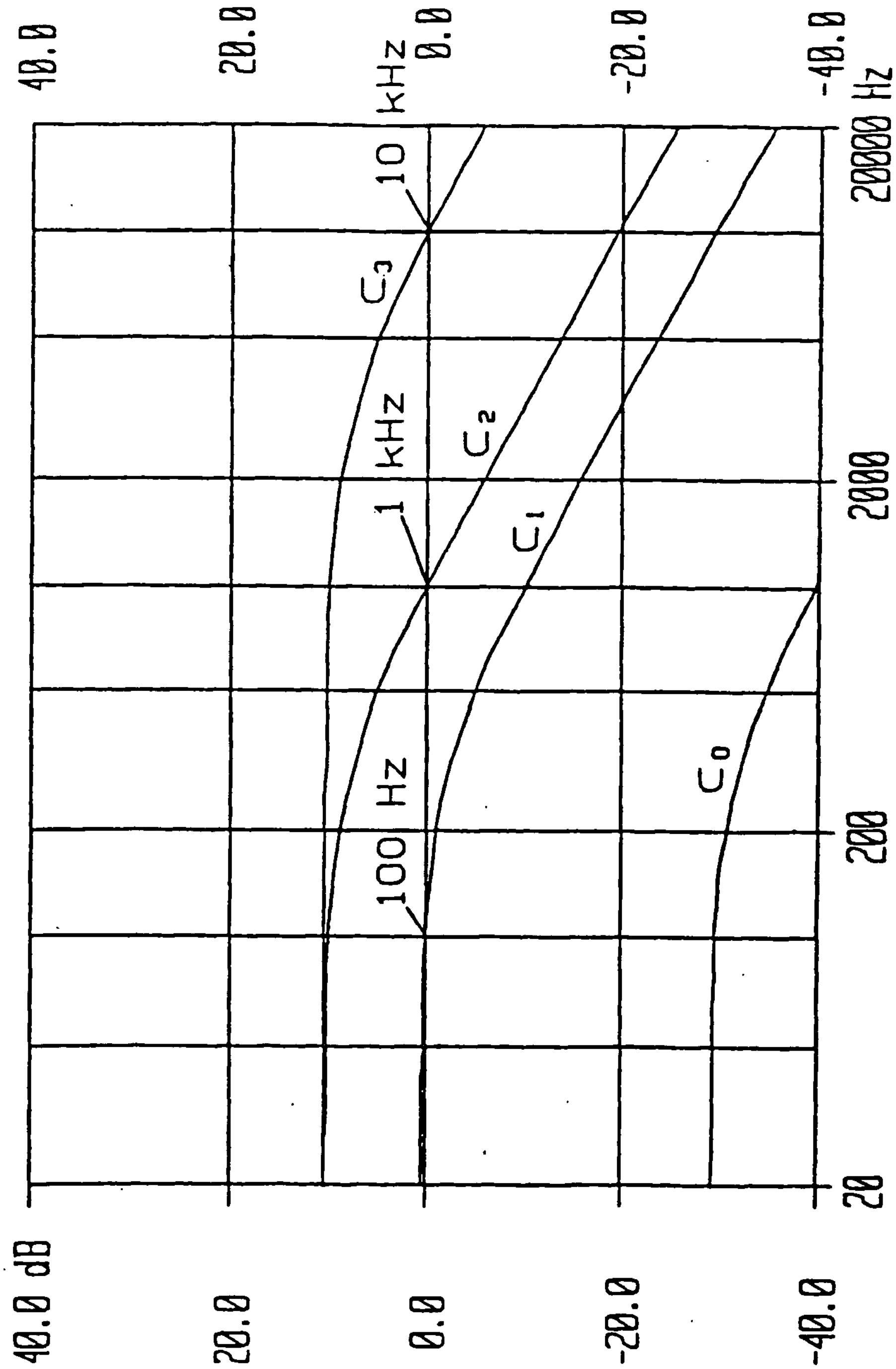
40.0 dB
20.0
0.0
-20.0
-40.0
20
200
2000
20000 Hz
100 Hz
1 kHz
10 kHz
C0
C1
C2
C3

Figure 6

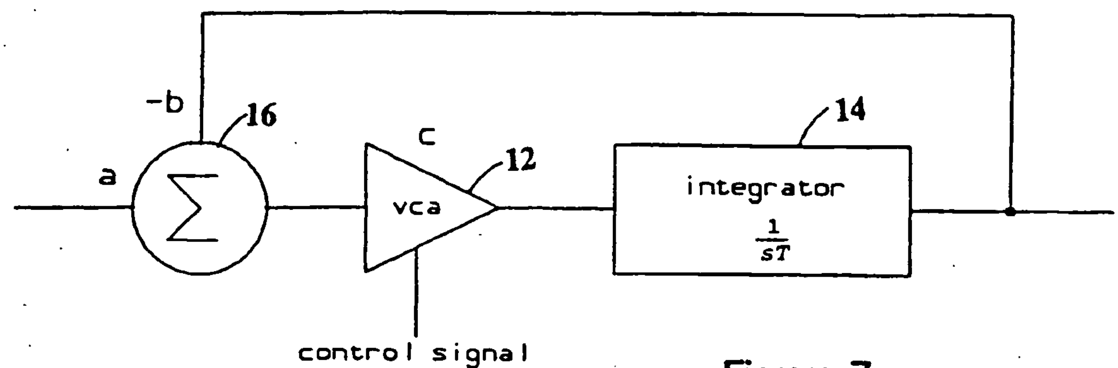


Figure 7
(Prior Art)

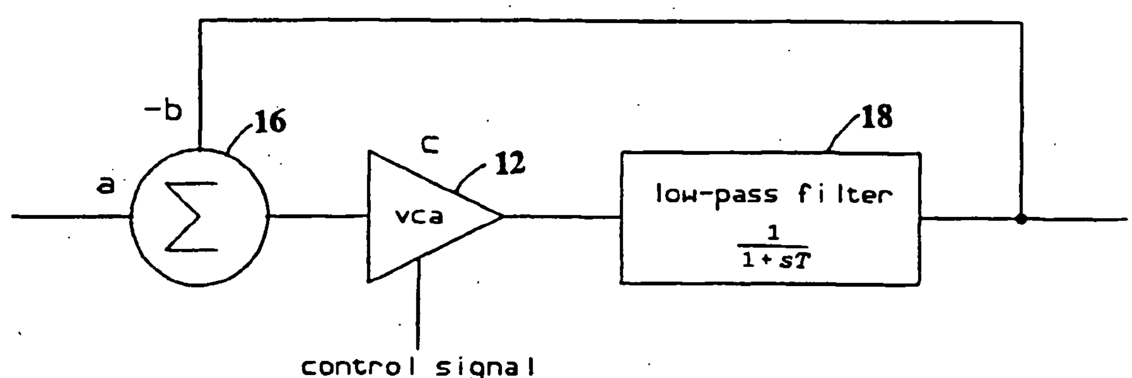


Figure 8

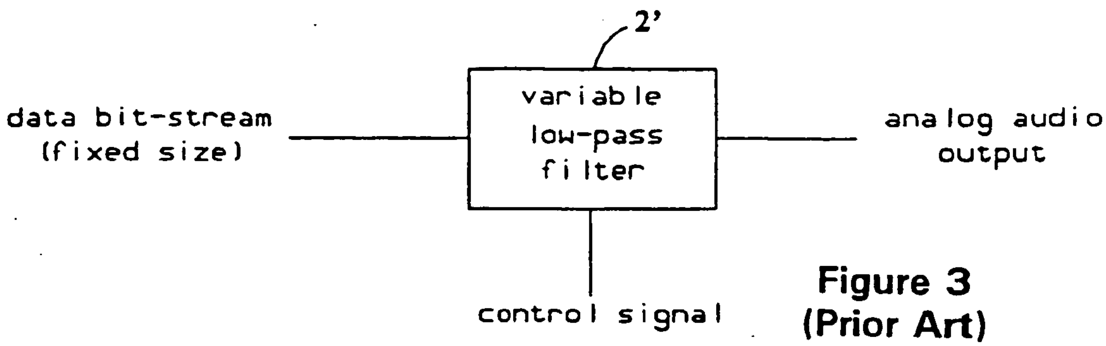


**Figure 3**
**(Prior Art)**

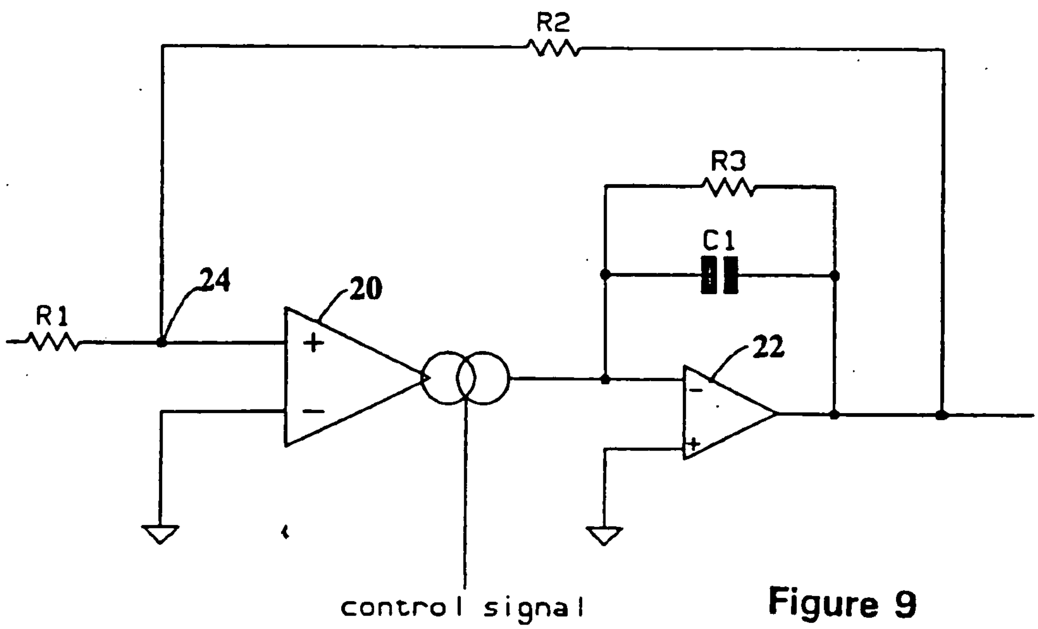


**Figure 9**

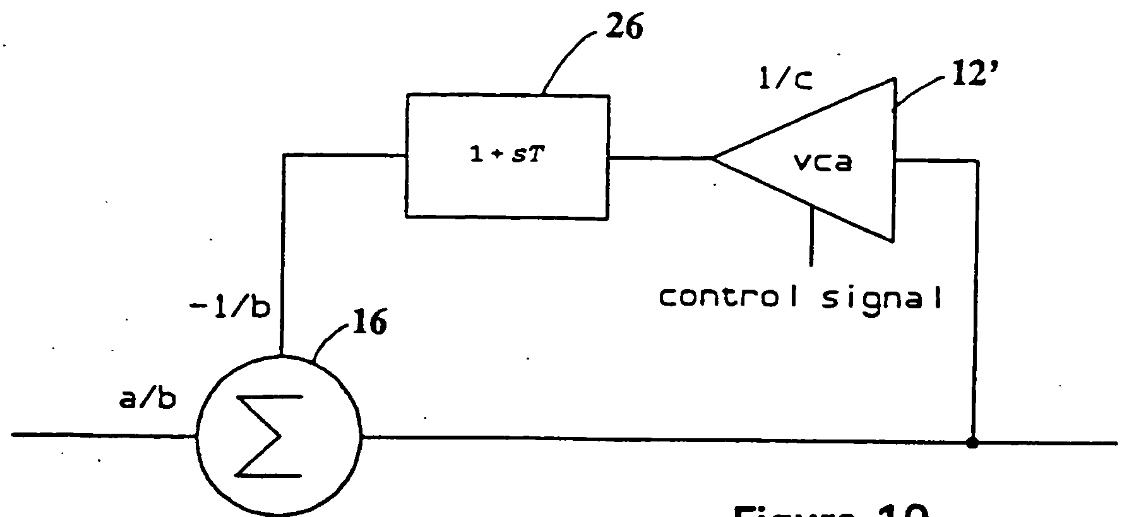


Figure 10

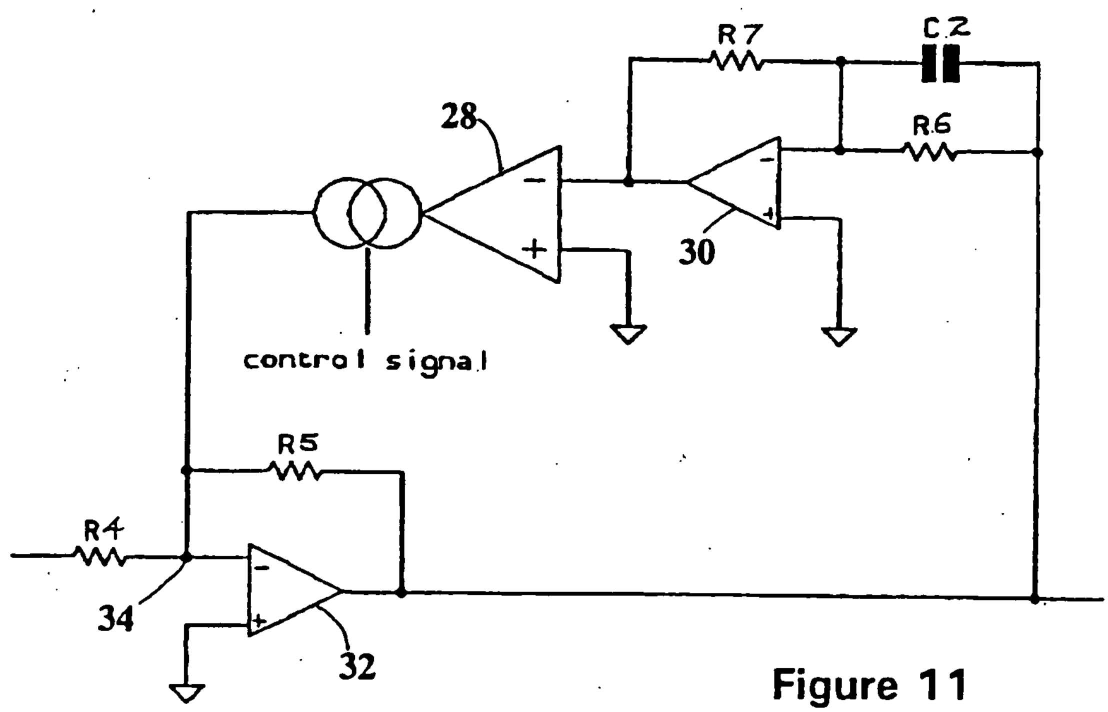


Figure 11